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REPORT ECOM-74-0272-8

ENVIRONMENT AND RADAR OPERATION SIMULATOR

By

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Engineering Experiment Station
Georgia Institute of Technology
Atlanta, Georgia 30332

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20. ABSTRACT (Continue on reverse side if necessary and identify by block number)

This report summarizes activities during the eighth quarter of a program sponsored by the U. S. Army Electronics Command (ECOM), directed toward the design of an Environment and Radar Operation Simulator (EROS). In addition some of the analysis results and design decisions are described in detail.

Algorithms for estimating the mean and variance of the filter outputs is described. The estimation takes into account the effects of multiplier truncation errors. The mean and variance are expressed in terms of the radar cross sections

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of the fixed and moving portions of simulated clutter, and the spectral parameters of the filter therefore impose limits on the magnitude of radar cross section that can be simulated. Tabulated results are presented.

PREFACE

This report was prepared at the Georgia Tech Engineering Experiment Station under Contract No. DAAB07-74-C-0272. The work covered by this report was performed in the Applied Engineering Laboratory under the supervision of Dr. H. A. Ecker and Mr. J. L. Eaves, Director of the Applied Engineering Laboratory and Chief of the Radar Technology Division, respectively. The progress reported herein was performed during the eighth quarter of a program to develop an environment and radar operation simulator (EROS) to be used in testing radar receivers and components.

This project is being monitored by Mr. Reinhard G. Olesch and Mr. Otto E. Rittenbach of the U. S. Army Electronics Command, and their helpful suggestions are acknowledged.

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1. INTRODUCTION

This report covers work performed during the period 1 April 1976 through 30 June 1976, the eighth quarter of a 30-month program to design and build an Environment and Radar Operation Simulation (EROS). The effort has been devoted primarily to continuing the implementation of EROS hardware and software.

A large part of the EROS hardware is completed through board design, wiring, and component assembly. The following major units have, in addition, been completed through unit test and debug:

<u>Computer Interface</u>

This unit moves data from the PDP-11 unibus to various locations in the EROS digital hardware. It provides protocol for conversing with PDP-11 I/O instructions and contains a small amount of buffer storage.

<u>1K by 16-bit Memory Module</u>

The 1K by 16-bit memory module is used in the clutter filter, the target processor, and in the video signal generator. The memory module includes drivers for addressing and enable lines.

<u>Clutter Timing and Control</u>

This unit generates the control sequence which defines the operations performed by the clutter hardware. It synchronizes the clutter and target processors, and it controls the filters, azimuth weighting, and azimuth integration.

<u>Clock Generator</u>

The clutter generation hardware receives 4 KHz and 10 MHz clock signals from the analog hardware. The 4 KHz clock is derived from the radar's P.R.F. The 10 MHz clock is derived from a crystal oscillator. The clock generator unit provides drivers of these signals, and it divides the 10 MHz clock by four yielding four phases.

Hardware Clutter Test

This unit provides facilities for troubleshooting and spot checking the operation of the clutter generation hardware. It is capable of forcing the clutter hardware to stop when any specified memory location is addressed, and it permits single-step operation from the stopped state. It also produces test data for exercising the clutter processor independent of the PDP-11 and of other EROS hardware.

A significant portion of the simulation preparation and real-time software has been completed. The following routines have been tested and debugged:

Spectral Clutter Library Maintenance

This collection of programs maintains a library of clutter spectral information. Each spectral record in the library is identified for reference purposes by an integer called a "spectral number". These programs diagnose spectral library updates to detect user-generated errors, derive filter statistics, and maintain a reference file in spectral-number order. The derivation of filter statistics is described in Section 2 of this report.

Complete Clutter Library Maintenance

This collection of programs maintains a library of complete clutter information. Each clutter record in the library contains spectral and radar cross-section information and is identified by a two-character name. These programs diagnose clutter updates to detect user-generated errors, derive filter parameters, and maintain a reference file of complete clutter information in identifier order.

Range-Dependent Reference Data

Each simulation run uses a table of 64 weights representing the range-dependent radar signal attenuations (e.g. R^{-4} power loss). These 64 weights are selected from a table of 197 entries corresponding to 60 m., 75.m . . . 3000 m. The particular entries selected depend upon the band of ranges being simulated. This program diagnoses the input data to detect user-generated errors and translates the data into 12-bit numbers for real-time simulations.

Target Recording

These programs collect (complex) samples of recorded target data from a dual-channel A/D converter and store the digitized data on PDP-11

disk for future reference. After the recording is finished the signal is reread from the disk to calculate its mean squared amplitude for the purpose of radar-cross-section scaling. Since the samples are read at a 4 KHz rate, one of the programs operates with real-time constraints.

Real-Time Executive

This program keeps track of time and antenna position. It invokes target processing of the illuminated target (if any), starts clutter input/output when appropriate, and permits the execution of any non-time-critical functions (e.g. CRT display)

Real-Time Target Processor

This program maintains target azimuth, obtains antenna weights from a core table, and transmits this data with target samples and range information to the EROS digital hardware. The real-time requirements for this program are critical, and the code is optimized for speed. The disk reading portion of this program (to fetch target data) has not yet been written, but the program has been tested and debugged with data initially generated in core.

2. CLUTTER FILTER TRUNCATION STATISTICS

The physical realization of the clutter-synthesis filter introduces errors due to the finite length of digital number representation. These errors are called truncation errors, and they occur specifically when two 16-bit numbers are multiplied to yield a 16-bit product. This section contains a brief review of the source of truncation errors followed by a description of the algorithms that compute their effect on the statistics of the filter outputs.

2.1 Source of Truncation Errors

Recall[1] that the complex digital filter which synthesizes clutter has two delay elements, whose outputs at time nT (after the nth iteration) are denoted by the complex quantities $w_1(nT)$ and $w_2(nT0$. The calculations performed by the filter are described by the equations

$$\begin{aligned} w_1\Big((n+1)T\Big) &= \alpha_1\, w_1\,(nT) + \alpha_2\, w_2(nT) + x(nT) \\ w_2\Big((n+1)T\Big) &= w_1\,(nT) \qquad\qquad (1) \\ v(nT) &= \beta\, w_1\,(nT) + K \end{aligned}$$

where

$x(nT)$ is the sequence of complex random inputs;

α_1 and α_2 are (real) parameters that control the clutter spectral statistics;

β is the (real) parameter that controls the radar cross section (RCS) of the moving portion of the clutter in the simulation cell;

K is the (complex) parameter that controls the RCS of the immobile portion of the clutter in the simulation cell;

$v(nT)$ is the sequence representing the sampled complex envelope of clutter return.

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[1] This formulation is extracted from Section 2 of the Sixth EROS Quarterly Report.

The filter is being implemented in the form of two identical modules; one generates real components of v(nT) for the 544 cells illuminated within an antenna beam, and the other generates imaginary components. For reasons discussed in Section 2.5 of the 5th EROS Quarterly Report, the product of $\alpha_1 w_1$ is actually formed as the sum $(\alpha - 1)w_1 + w_1$. The numbers in the digital filter modules are represented as 16-bit 2's complement binary fractions with the assumed binary point between the sign bit and 15 fraction bits. Thus during every computation cycle the real and imaginary components of the two products $(\alpha_1 - 1)w_1$ and $\alpha_2 w_2$ are truncated from 30 fraction bits to 15 fraction bits.

The errors introduced by the truncation are expressed as follows. For any real number a, let [a] denote the largest integer that does not exceed a. For any complex number a + jb, let [a + jb] denote [a] + j[b]. Then the truncated products are $[2^{15}(\alpha_1 - 1)w_1]2^{-15}$ and $[2^{15}\alpha_2 w_2]2^{-15}$, and the truncation errors are

$$e_1 = [2^{15}(\alpha_1 - 1)w_1]\, 2^{-15} - (\alpha_1 - 1)w_1$$

and (2)

$$e_2 = [2^{15}\alpha_2 w_2]2^{-15} - \alpha_2 w_2 \quad .$$

The effects of truncation errors can be analyzed by treating e_1 and e_2 as additional sources of random input, because Equation (1) is still valid if x(nT) is considered to be the sum

$$x(nT) = e_1(nT) + e_2(nT) + r(nT) \tag{3}$$

where r(nT) is the output produced by the pseudo-random number generator.[1]

2.2 Relevance of Truncation Statistics

For any random variable u, let $\mu\{u\}$ and $s^2\{u\}$ denote its mean and variance respectively. If u is complex with real part u_R and imaginary part u_I, then

$$\mu\{u\} = \mu\{u_R\} + j\mu\{u_I\}$$

and (4)

$$s^2\{u\} = s^2\{u_R\} + s^2\{u_I\}$$

[1]See Section 2.3 of the 6th EROS Quarterly Report.

Using this notation, the third equation in (1) can be restated in the form

$$v(nT) = \beta[w_1(nT) - \mu\{w_1\}] + \beta\mu\{w_1\} + K . \quad (5)$$

The first term in (5) has zero mean and represents the radar return from the moving scatterers in the clutter cell. The second term and third term in Equation (5) are constant, and their sum represents the radar return from immobile scatterers in the clutter cell. The RCS of the moving scatterers is, therefore proportional[1] to the mean squared amplitude of $\beta[w_1(nT) - \mu\{w_1\}]$, which equals $\beta^2 s^2\{w_1\}$. Similarly the RCS of the immobile scatterers is proportional to $|\beta\mu\{w_1\} + K|^2$. Thus in order to calculate β and K from the moving and immobile RCS values, it is necessary to estimate $\mu\{w_1\}$ and $s^2\{w_1\}$. These quantities, in turn, can be computed from the following formulas.[2]

$$F = \frac{1}{1 - \alpha_1 - \alpha_2}$$

$$G^2 = \frac{1 - \alpha_2}{(1 + \alpha_2)[(1 - \alpha_2)^2 - \alpha_1^2]} \quad (6)$$

$$\mu\{w_1\} = F[(1 + j)\,\mu\{\rho\} + \mu\{e_1\} + \mu\{e_2\}]$$

$$s^2\{w_1\} = F^2(-2\gamma\mu\{\rho\} - 2[\mu\{\rho\}]^2) + G^2(2\gamma^2 + s^2\{e_1\} + s^2\{e_2\} + 2\gamma\mu\{\rho\})$$

The parameter γ is the magnitude of the pseudo random input and is selected as the largest number from the set $\{2^{-1}, 2^{-2}, \ldots, 2^{-15}\}$ for which the filter is safe from overflow (i.e. for which $|\mathrm{Re}\{w_1(nT)\}| < 1$ or $|\mathrm{Im}\{w_1(nT)\}| < 1$). The quantity $\mu\{\rho\}$ equals $\gamma(2^{31} - 1)^{-1}$. From these relationships it follows that the calculation of β and K from the two clutter RCS values requires that γ, $\mu\{e_1\}$, $s^2\{e_1\}$, $\mu\{e_2\}$, and $s^2\{e_2\}$ be evaluated.

It is clear from symmetry that for each of the random variables $x(nT)$, $e_1(nT)$, $e_2(nT)$, and $w_1(nT)$, any property which holds for the real

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[1] Refer to Section 3 of the 7th EROS Quarterly Report.

[2] Refer to Section 2.3 of the 6th EROS Quarterly Report.

part of the random variable also holds for the imaginary part. Let the subscript R denote "the real part of" and the subscript I denote "the imaginary part of". Then for any of the above random variables u

$$\mu\{u\} = (1 + j)\,\mu\{u_R\} = (1 + j)\,\mu\{u_I\}$$
$$s^2\{u\} = 2s^2\{u_R\} = 2s^2\{u_I\} \,. \tag{7}$$

2.3 Evaluation of γ

To calculate γ the following inequality is employed:[1]

$$|\, w_{1R}(nT) \,| < |\, \hat{x}_R \,| \sum_{n=1}^{\infty} |\, h(nT) \,| \,, \tag{8}$$

where $|\,\hat{x}_R\,|$ is the maximum of $|\,x_R\,(nT)\,|$ and $h(nT)$ is the filter's impulse response. Let $\hat{e}_{1R}$ and $\hat{e}_{2R}$ denote maximum values of $|\,e_{1R}\,(nT)\,|$ and $|\,e_{2R}\,(nT)\,|$ respectively. Then clearly if

$$(\gamma + \hat{e}_{1R} + \hat{e}_{2R}) \sum_{n=1}^{\infty} |\, h(nT) \,| < 1 \,. \tag{9}$$

the filter will not overflow with pseudo-random input of magnitude γ.

The values of e_{1R} and e_{2R} are easily calculated from

$$\hat{e}_{1R} = 2^{-15}\,(1 - 2^{-m_1})$$
$$\hat{e}_{2R} = 2^{-15}\,(1 - 2^{-m_2}) \tag{10}$$

where m_1 and m_2 are the smallest respective integers such that $2^{m_1}(\alpha_1 - 1)$

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[1] c. f. Oppenheim, A. V., and Schafer, R. W., Digital Signal Processing, Prentice-Hall, Englewood Cliffs, N. J., 1975, p. 427.

and $2^{m_2}\alpha_2$ are integers. The evaluation of $\sum_{n=1}^{\infty} | h(nT) |$ is somewhat more complicated. The denominator of the z-transform of h(nT) is $z^2 - \alpha_1 z - \alpha_2$, and if the polynomial either has two identical roots or has 0 for one of its roots, then a closed form expression for $\sum_{n=1}^{\infty} | h(nT) |$ can be derived. In the two identical roots case

$$\sum_{n=1}^{\infty} | h(nT) | = \frac{1}{\left(1 - \frac{|\alpha_1|}{2}\right)^2} \quad \text{when } \alpha_1^2 + 4\alpha_2 = 0 \qquad (11)$$

In the 0-root case

$$\sum^{\infty} |h(nT)| = \frac{1}{1 - |\alpha_1|} \quad \text{when } \alpha_2 = 0 \qquad (12)$$

Although no closed-form expression has been found for the general case, a numerical evaluation of the sum is straight forward. To approximate the sum the following expression for a bound on the remainder has been found useful

$$\sum_{n=1}^{\infty} | h(nT) | - \sum_{n=1}^{m} | h(nT) | < \frac{2\, a_{max}^{m+1}}{(1 - a_{max})\sqrt{|\alpha_1^2 + 4\alpha_2|}} \quad (\alpha_1^2 + 4\alpha_2 \neq 0) \qquad (13)$$

where a_{max} is the magnitude of the larger of the two roots:

$$a_{max} = \frac{1}{2} \max \left\{ \left| \alpha_1 \pm \sqrt{\alpha_1^2 + 4\alpha_2} \right| \right\} \qquad (14)$$

A computer program to calculate γ using these formulas has been written and debugged.

2.4 Evaluation of Truncation-Error Means and Variances

An iterative approximation method is being employed to estimate the means and variances of e_1 and e_2. Observe that Equation (2) expresses the random variables e_1 and e_2 as functions of w_1 and w_2, where $\alpha_1 - 1$ and α_2 are constants. We define

$$e_\alpha (w) = \left[2^{15} \alpha w \right] 2^{-15} - \alpha w , \qquad (15)$$

so that $e_1 = e_{\alpha_1 - 1} (w_1)$ and $e_2 = e_{\alpha_2} (w_2)$. Since the only difference between w_1 and w_2 is a delay of one step, the probability distributions of w_{1R} and w_{2R} are identical. Therefore, if their density function $P(w_{1R})$ were known, then the means and variances of the error variables could be computed from

$$\mu \left\{ e_\alpha (w_{1R}) \right\} = \sum_{w_{1R}} e_\alpha (w_{1R}) P (w_{1R}) \qquad (16)$$

and

$$s^2 \left\{ e_\alpha (w_{1R}) \right\} = \sum_{w_{1R}} e_\alpha^2 (w_{1R}) P (w_{1R}) - \left[\mu \left\{ e_\alpha (w_{1R}) \right\} \right]^2 \qquad (17)$$

It is necessary to treat $P(w_{1R})$ as a discrete probability distribution function (and to use sums instead of integrals), because for most of the values of α, $e_\alpha (w_{1R})$ is not even approximately a continuous function of w_{1R}. The discrete set of values assumed by w_{1R} is $\{\pm i 2^{-15}: i = 0, \ldots 2^{15} - 1\}$.

Unfortunately $P(w_{1R})$ is not known. However, computer simulations have empirically shown that the distribution of w_{1R} is approximately Gaussian[1]. Therefore, we can assume the approximation

$$P(i2^{-15}) \cong \frac{2^{-15}}{\sqrt{2\pi}\; s\{w_{1R}\}} \exp \frac{(i2^{-15} - \mu\{w_{1R}\})^2}{2s^2\{w_{1R}\}} \qquad (18)$$

---

[1] Refer to Section 2.4 of the Fifth EROS Quarterly Report.

and iteratively estimate $\mu\{w_{1R}\}$ and $s^2\{w_{1R}\}$. The iterative algorithm is more specifically the following:

1. Initialize $\mu\{e_{1R}\}$ and $\mu\{e_{2R}\}$ to $\frac{1}{2}\hat{e}_{1R}$ and $\frac{1}{2}\hat{e}_{2R}$ respectively. (Refer to Equation (10). The set of possible values assumed by e_1 is $\left\{0,\ 1 \cdot 2^{-15-m_1},\ \ldots,\ (2^{m_1}-1)2^{-15-m_1}\right\}$. The largest value in this set is $\hat{e}_{1R}$, and the average value--assuming equal likelihood for each possibility -- is $\frac{1}{2}\hat{e}_{1R}$. The same discussion applies to $\hat{e}_2$.) Initialize $s^2\{e_{1R}\}$ and $s^2\{e_{2R}\}$ to zero.

2. Calculate $\mu\{w_{1R}\}$ and $s^2\{w_{1R}\}$ using (6).

3. Recalculate $\mu\{e_{1R}\}$ and $s^2\{e_{1R}\}$ using (15), (16), (17) and (18).

4. A heuristic criterion for convergence is applied: another iteration starting at Step 2 is performed if the newly computed values of $\mu\{e_{1R}\}$ and $s^2\{e_{1R}\}$ are not sufficiently close to the previous estimates.

The computation of $e_\alpha(w_{1R})$ is somewhat easier than that suggested by Equation (15). The following algorithm is employed in the program which calculates means and variances.

1. Compute r_1, residue (modulo 2^{15}) [1] of $2^{15}\alpha$.

2. Compute r_2, the residue (modulo 2^{15}) of w_{1R}.

3. Compute r_3, the residue (modulo 2^{15}) of $r_1 r_2$.

4. $e_\alpha(w_{1R}) = 2^{-30} r_3$

---

[1] For any two integers m and x, the residue (modulo m) of x is the integer r such that $x = qm + r$, where q is an integer and $0 \leq r < m$.

3. MAXIMUM RADAR CROSS SECTION FOR CLUTTER

3.1 Introduction

The radar cross section corresponding to an EROS signal is proportional to its average squared amplitude.[1] More specifically

$$\sigma = K_\sigma \overline{A^2}$$

where

σ is the RCS of the simulated scatterers,

K_σ is the proportionality constant (20,000 m^2 for AN/PPS-15 simulations), and

$\overline{A^2}$ is the average squared amplitude.

The components of RCS associated with a given clutter cell are denoted by σ_R and σ_S for the moving and immobile scatterers respectively. Since their respective mean squared amplitudes are $\beta^2 s^2\{w_1\}$ and $| K + \beta\mu\{w_1\} |^2$ (refer to the previous section of this report),

$$\begin{aligned} \sigma_R &= K_\sigma \beta^2 s^2\{w_1\} = 2K_\sigma \beta^2 s^2\{w_{1R}\} \\ \sigma_S &= K_\sigma |K + \beta\mu\{w_1\}|^2 = K_\sigma |K + \beta\mu\{w_{1R}\}(1 + j)|^2 \end{aligned} \tag{20}$$

3.2 Parameter Constraints

The magnitude of the parameter β has an upper bound ν, which is determined by the position of the assumed binary point in the digital hardware. If there are n bits between the sign and the assumed binary point, then $\nu = 2^n$. In the current implementation n = 0; therefore, $\nu = 1$. Since σ_R is proportional to β^2 as expressed by (20),

---

[1] Refer to the Section 3 of the Seventh EROS Quarterly Report.

$$\sigma_R < 2K_\sigma \nu^2 s^2 \{w_{1R}\} \tag{21}$$

The real and imaginary components of K are also limited by the positions of the assumed binary point. This binary point position has been selected to be consistent with that of β. Thus

$$\begin{aligned} | K_R | &< \nu \\ | K_I | &< \nu \end{aligned} \tag{22}$$

These inequalities lead to a mutual constraint between σ_R and σ_S.

The computer program CCDIAB maintains the library of reference data for different types of clutter. The user supplies spectral and RCS information, and CCDIAB stores this data in a file for subsequent retrieval. CCDIAB calculates β from σ_R using the first formula in (20).

$$\beta = \sqrt{\frac{\sigma_R}{2K_\sigma s^2 \{w_{1R}\}}} \tag{23}$$

However CCDIAB does not have sufficient information to calculate K in similar fashion because the phase of K is not determined until scenario compilation time. The clutter scenario compilation program, which utilizes the reference file produced by CCDIAB, randomly assigns the phase of $S = K + \beta\mu\{w_1\}$ from a uniform distribution between 0° and 360°, and assigns $\sqrt{\sigma_S/K_\sigma}$ to the magnitude of S in accordance with (20). Then with S defined, K can be calculated

$$K = S - \beta\mu\{w_1\} \tag{24}$$

In order to satisfy (22)

$$|S_R - \beta\mu\{w_{1R}\}| < \nu$$

and (25)

$$|S_I - \beta\mu\{w_{1R}\}| < \nu$$

Utilizing the fact that $|\mu\{w_{1R}\}| < 1$ and that $\beta < \nu$, it can be shown that the inequalities (25) are true for all complex numbers S of magnitude $\sqrt{\sigma_S/K_\sigma}$ if and only if

$$\sigma_S < K_\sigma \left(\nu - \beta \, |\mu\{w_{1R}\}|\right)^2 \tag{26}$$

Combining (23) and (26)

$$\sigma_S < K_\sigma \left(\nu - \left|\mu\left\{w_{1R}\right\}\right| \sqrt{\frac{\sigma_R}{2K \; s^2\left\{w_{1R}\right\}}}\right)^2 \tag{27}$$

Inequality (27) expresses a mutual constraint limiting the magnitude of the radar cross sections σ_R and σ_S.

The following is a summary of the computations and tests performed by CCDIAB, which ensure that the pertinent constraints are met.

1. The radar cross section σ_R of the moving scatterers is compared to the maximum value $2K_\sigma \nu^2 s^2\{w_{1R}\}$ implied by (21). If σ_R is too large, it is reduced to the maximum value.

2. β is calculated from σ_R using (23).

3. The radar cross section σ_S of the immobile scatterers is compared to the maximum value $K_\sigma (\nu - \beta|\mu\{w_{1R}\}|)^2$ implied by (26). If σ_S is too large it is reduced to the maximum value.

3.3 Dependence of Maximum RCS on Spectral Parameters

The radar cross section of the immobile clutter scatterers is, for all practical purposes, non-constrained. Values of $|\mu\{w_{1R}\}|$ tend to be very small (much less than .5). Therefore, the maximum value of σ_S that satisfies (26) is greater than 5000 m^2 for worst-case β ($\beta = 1$), and 5000 m^2 is only about 6 dB below the largest EROS RCS (20,000 m^2).

The radar cross section of the moving scatterers is, on the other hand, sensitive to the spectral parameters. Maximum values of σ_R for typical cutoff frequencies (f_c), calculated from (21), are tabulated in Table I.

TABLE I. Maximum Radar Cross Section (m^2)

Cutoff Frequency f_c	σ_R max
20	211
15	124
10	100
5	12

Values of $s^2 \{w_{1R}\}$ were estimated by the computer program which uses the algorithm described in Section 2. The spectral parameters α_1 and α_2 in these calculations were derived from two-pole analog-filters whose s-plane poles subtend $\pm$ 150° with the positive real axis and have magnitudes appropriate for the given cutoff frequency.

Table I illustrates the fact that $s^2\{w_1\}$ decreases quickly with f_c. This is not surprising since the dominant term in the Formula (6) for $s^2\{w_1\}$ is $2G^2 \gamma^2$, and

$$G^2 \gamma^2 \doteq \frac{\Sigma h^2 (nT)}{(\Sigma \mid h(nT) \mid)^2} \qquad (28)$$

(the numerator equals G^2 by definition). As f_c decreases, the number of non-negligible items in h(nT) increases, and the ratio in (28) is therefore reduced.

4. NEXT QUARTER PLANS

The results presented in Section 2 and 3 represent the completion of the analysis to support EROS design. During the next quarter the effort will be devoted primarily to implementation, and design additions or modifications will be performed only if omissions or errors are identified. It is anticipated that all of the EROS hardware will have been completed by the end of next quarter through unit assembly and test. As time permits part of the hardware integration may be performed. The plans for software development next quarter include clutter scenario definition via CRT, target scenario definition via CRT, clutter scenario compilation, and real-time clutter processing.

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